DESCRIPTION

WIRELESS TERMINAL

The present invention relates to a wireless terminal, for example a mobile phone handset.

Wireless terminals, such as mobile phone handsets, typically incorporate either an external antenna, such as a normal mode helix or meander line antenna, or an internal antenna, such as a Planar Inverted-F Antenna (PIFA) or similar.

Such antennas are small (relative to a wavelength) and therefore, owing to the fundamental limits of small antennas, narrowband. However, cellular radio communication systems typically have a fractional bandwidth of 10% or more. To achieve such a bandwidth from a PIFA for example requires a considerable volume, there being a direct relationship between the bandwidth of a patch antenna and its volume, but such a volume is not readily available with the current trends towards small handsets. Hence, because of the limits referred to above, it is not feasible to achieve efficient wideband radiation from small antennas in present-day wireless terminals.

A further problem with known antenna arrangements for wireless terminals is that they are generally unbalanced, and therefore couple strongly to the terminal case. As a result a significant amount of radiation emanates from the terminal itself rather than the antenna. A wireless terminal in which an antenna feed is directly coupled to the terminal case, thereby taking advantage of this situation, is disclosed in our co-pending unpublished International patent application PCT/EPO1/08550 (Applicant's reference PHGB010056). When fed via an appropriate matching network the terminal case acts as an efficient, wideband radiator.

An object of the present invention is to provide a compact wireless terminal having efficient radiation properties without the need for a matching network.

According to the present invention there is provided a wireless terminal comprising a ground conductor and a transceiver coupled to an antenna feed, wherein the antenna feed is coupled directly to the ground conductor via a capacitor formed by a conducting plate and a portion of the ground conductor and wherein a slot, partially located underneath the conducting plate, is provided in the ground conductor.

The location of a slot beneath the conducting plate performs much of the function of a conventional matching circuit, thereby simplifying implementation of a wireless terminal. More than one slot may be provided, and a slot may be folded as dictated by space or other requirements.

The present invention is applicable to any wireless communication system where the use of a large antenna is not appropriate. Since the coupling capacitor is small, it is ideally suited to an RF IC or module, where the coupling capacitor would be part of the module. It is particularly useful in wireless systems that feature multiband or wideband operation.

The present invention is based upon the recognition, not present in the prior art, that the impedances of an antenna and a wireless handset are similar to those of an asymmetric dipole, which are separable, and on the further recognition that the antenna impedance can be replaced with a non-radiating coupling element.

Embodiments of the present invention will now be described, by way of example, with reference to the accompanying drawings, wherein:

Figure 1 shows a model of an asymmetrical dipole antenna, representing the combination of an antenna and a wireless terminal;

Figure 2 is a graph demonstrating the separability of the components of the impedance of an asymmetrical dipole;

Figure 3 is an equivalent circuit of the combination of a handset and an antenna;

Figure 4 is an equivalent circuit of a capacitively back-coupled handset;

Figure 5 is a perspective view of a basic capacitively back-coupled handset;

Figure 6 is a graph of simulated return loss S_{11} in dB against frequency f in MHz for the handset of Figure 5;

Figure 7 is a Smith chart showing the simulated impedance of the handset of Figure 5 over the frequency range 1000 to 2800MHz;

Figure 8 is a graph showing the simulated resistance of the handset of Figure 5;

Figure 9 is a plan view of a single-slotted self-resonant capacitively back-coupled handset;

Figure 10 is a graph of simulated return loss S_{11} in dB against frequency f in MHz for the handset of Figure 9;

Figure 11 is a Smith chart showing the simulated impedance of the handset of Figure 9 over the frequency range 800 to 3000MHz;

Figure 12 is a plan view of a doubly-slotted self-resonant capacitively back-coupled handset;

Figure 13 is a graph of simulated return loss S_{11} in dB against frequency f in MHz for the handset of Figure 12;

Figure 14 is a Smith chart showing the simulated impedance of the handset of Figure 12, over the frequency range 800 to 3000MHz;

Figure 15 is a graph of simulated return loss S_{11} in dB against frequency f in MHz for the handset of Figure 12 fed via a matching network; and

Figure 16 is a Smith chart showing the simulated impedance of the handset of Figure 12 fed via a matching network, over the frequency range 800 to 3000MHz.

In the drawings the same reference numerals have been used to indicate corresponding features.

Figure 1 shows a model of the impedance seen by a transceiver, in transmit mode, in a wireless handset at its antenna feed point. The impedance is modelled as an asymmetrical dipole, where the first arm 102 represents the impedance of the antenna and the second arm 104 the impedance of the handset, both arms being driven by a source 106. As shown in the figure, the impedance of such an arrangement is substantially equivalent to the sum of the impedance of each arm 102,104 driven separately against a virtual ground 108. The model could equally well be used for reception by replacing the source 106 by an impedance representing that of the transceiver, although this is rather more difficult to simulate.

The validity of this model was checked by simulations using the well-known NEC (Numerical Electromagnetics Code) with the first arm 102 having a length of 40mm and a diameter of 1mm and the second arm 104 having a length of 80mm and a diameter of 1mm. Figure 2 shows the results for the real and imaginary parts of the impedance (R+jX) of the combined arrangement (Ref R and Ref X) together with results obtained by simulating the impedances separately and summing the result. It can be seen that the results of the simulations are quite close. The only significant deviation is in the region of half-wave resonance, when the impedance is difficult to simulate accurately.

An equivalent circuit for the combination of an antenna and a handset, as seen from the antenna feed point, is shown in Figure 3. R_1 and jX_1 represent the impedance of the antenna, while R_2 and jX_2 represent the impedance of the handset. From this equivalent circuit it can be deduced that the ratio of power radiated by the antenna, P_1, and the handset, P_2, is given by

$$\frac{P_1}{P_2} = \frac{R_1}{R_2}$$

If the size of the antenna is reduced, its radiation resistance R_1 will also reduce. If the antenna becomes infinitesimally small its radiation resistance R_1 will fall to zero and all of the radiation will come from the handset. This situation can be made beneficial if the handset impedance is suitable for the source 106 driving it and if the capacitive reactance of the infinitesimal

antenna can be minimised by increasing the capacitive back-coupling to the handset.

With these modifications, the equivalent circuit is modified to that shown in Figure 4. The antenna has therefore been replaced with a physically very small back-coupling capacitor, designed to have a large capacitance for maximum coupling and minimum reactance. The residual reactance of the back-coupling capacitor can be tuned out with a simple matching circuit. By correct design of the handset, the resulting bandwidth can be much greater than with a conventional antenna and handset combination, because the handset acts as a low Q radiating element (simulations show that a typical Q is around 1), whereas conventional antennas typically have a Q of around 50.

A basic embodiment of a capacitively back-coupled handset is shown in Figure 5. A handset 502 has dimensions of 10×40×100mm, typical of modern cellular handsets. A parallel plate capacitor 504, having dimensions 2×10×10mm, is formed by mounting a 10×10mm plate 506 2mm above the top edge 508 of the handset 502, in the position normally occupied by a much larger antenna. The resultant capacitance is about 0.5pF, representing a compromise between capacitance (which would be increased by reducing the separation of the handset 502 and plate 506) and coupling effectiveness (which depends on the separation of the handset 502 and plate 506). The capacitor is fed via a support 510, which is insulated from the handset case 502.

The return loss S_{11} of this embodiment after matching was simulated using the High Frequency Structure Simulator (HFSS), available from Ansoft Corporation, with the results shown in Figure 6 for frequencies f between 1000 and 2800MHz. A conventional two inductor “L” network was used to match at 1900MHz. The resultant bandwidth at 7dB return loss (corresponding to approximately 90% of input power radiated) is approximately 60MHz, or 3%, which is useful but not as large as was required. A Smith chart illustrating the simulated impedance of this embodiment over the same frequency range is shown in Figure 7.

The low bandwidth is because the combination of the handset 502 and capacitor 504 present an impedance of approximately 3−j90Ω at 1900MHz. Figure 8 shows the resistance variation, over the same frequency range as before, simulated using HFSS. This can be improved by redesigning the case to increase the resistance, for example by the use of a slot or a narrower handset, as discussed in our co-pending unpublished International patent application PCT/EPO1/08550.

The handset of Figure 5 requires matching to obtain reasonable performance. There are significant advantages to being able to eliminate the need for matching. A plan view of a modified single band configuration which requires no matching is shown in Figure 9. This embodiment differs from that of Figure 5 in that the 10mm square plate 506 is located 2mm above the back of the handset 502, and in that a slot 912 of length 30mm and width 1mm is cut in the conducting material 2mm from the edge of the handset case. The slot 912 extends under the conducting plate 506 (as shown by dashed lines in Figure 9). The slot 912 is resonant at odd multiples of a quarter wavelength, i.e. at $\lambda/4$, 3 $\lambda/4$, etc.

The slot presents a high impedance to the coupling capacitor, thereby enabling a good match to 50Ω. It is believed that the capacitor excites a transmission line mode in the slot 912 that acts as a shunt inductance at the antenna feed, which acts to match the response.

In the illustrated embodiment the slot 912 is located close to the edge of the handset case 502 in order to minimise the space used, although the slot could equally well be located on the other side of the coupling capacitor 504. Similarly, the coupling capacitor could be implemented in other positions on the handset 502 and the slot 912 could have a range of configurations, for example vertical, horizontal or meandering.

The return loss S_{11} of this embodiment, without matching, was simulated using HFSS, with the results shown in Figure 10 for frequencies f between 800 and 3000MHz. The resultant bandwidth at 7dB return loss is approximately 90MHz, or 4.3%. Although the bandwidth could be improved with matching, it is useful to be able to avoid having to include matching and

the bandwidth is already more than sufficient for a Bluetooth embodiment, for example.

A Smith chart illustrating the simulated impedance of this embodiment over the same frequency range is shown in Figure 11. This shows that the configuration of Figure 9 also has the useful property that resonance (zero reactance) is achieved twice, with the higher frequency resonance having the higher resistance. This is particularly convenient, since the receive band is usually at a higher frequency in a frequency duplex system.

A preferred transceiver architecture is to maintain a low impedance path between the (generally low impedance) transmitter and the antenna, and a high impedance path between the antenna and the (generally high impedance) receiver. However, for simplicity of design it is conventional to use a 50Ω system impedance with additional matching at the transmitter and receiver as required. This matching is lossy, and may also reduce the bandwidth seen at both the transmitter and receiver. Hence, the removal of the need for matching is a significant advantage of the present invention.

A dual band embodiment of the present invention is shown in plan view in Figure 12. In this embodiment the plate 506 and slot 912 have been moved to the top centre of the back surface of the handset 502, and a further slot 1214 has been added. The further slot 1214 is longer than the first slot 912, having a a total length of approximately 73mm and a width of 1mm, and folded to reduce the area it occupies.

The return loss S_{11} of this embodiment, without matching, was simulated using HFSS, with the results shown in Figure 13 for frequencies f between 800 and 3000MHz. It can clearly be seen that this design allows dual, tri or multiband operation. The slots 912, 1214 are resonant at odd multiples of $\lambda/4$, and can therefore be arranged to give individual or combined resonances. The first resonance (at approximately 1GHz) is the $\lambda/4$ resonance of the longer slot 1214. The second resonance (at approximately 1.8GHz) is the $\lambda/4$ resonance of the shorter slot 912. The third resonance (at approximately 2.8GHz) is the 3 $\lambda/4$ resonance of the longer slot 1214. It is clear, for example,

that, with some modification, this configuration can be used for GSM, DCS1800 and Bluetooth.

The resultant bandwidths at 7dB return loss for the three resonances are approximately 15MHz (1.5%), 110MHz (5.9%) and 110MHz (3.9%). The bandwidth of the 1GHz resonance is small, but the other bandwidths are good. A Smith chart illustrating the simulated impedance of this embodiment over the same frequency range is shown in Figure 13. The rapid changes in impedance in the Smith chart reflect the narrow-band nature of the first resonance.

The self-resonance of each slot 912,1214 is independently variable via its position under the feeding capacitor 504: as the slot 912,1214 is progressively moved under the plate 506 the effect of its nominal shunt inductance increases. Also, each slot 912,1214 is high impedance at its open end and low impedance at its shorted end. Hence, the resistance could be varied by tapping off at various points along the slot. The capacitor can also be made asymmetric to allow for such tapping to be performed, to some extent.

Embodiments of the present invention may also be used in conjunction with matching. As an example, simulations of the dual slot configuration illustrated in Figure 12 in conjunction with a simple "L" matching circuit similar to that used for the basic embodiment of Figure 5 were performed. Results for the return loss S_{11} are shown in Figure 15 for frequencies f between 800 and 3000MHz. It can be seen that a very wide bandwidth is achieved (a 3dB bandwidth of approximately 1.4GHz). This could be enhanced further with a more elaborate matching circuit. A Smith chart illustrating the simulated impedance of this embodiment over the same frequency range is shown in Figure 16.

In the above embodiments a conducting handset case has been the radiating element. However, other ground conductors in a wireless terminal could perform a similar function. Examples include conductors used for EMC shielding and an area of Printed Circuit Board (PCB) metallisation, for example a ground plane.

From reading the present disclosure, other modifications will be apparent to persons skilled in the art. Such modifications may involve other

features which are already known in the design, manufacture and use of wireless terminals and component parts thereof, and which may be used instead of or in addition to features already described herein.